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Thesis

THE TRANSIENT SIMULATION OF OPTOELECTRONIC INTEGRATED CIRCUITS USING ‘SPICE’

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ABSTRACT

A new transient simulation technique for optoelectronic devices and circuits is demonstrated using a novel approach for incorporating the solution of semiconductor drift-diffusion equations within a general purpose circuit simulation program (SPICE). In this approach, the time dependent semiconductor models, namely Poisson’s and drift-diffusion equations are discretized on a space grid and are represented by lumped circuit elements such as nonlinear dependent sources, capacitors, and resistors. Accurate transient simulation of high-speed optoelectronic devices under realistic bias conditions is becoming an increasingly important concern in Optoelectronic Integrated Circuit (OEIC) design and optimization. For the development of optoelectronic devices and circuits, a simulation tool capable of handling physical phenomena such as heterojunctions, large signal effects, local optical generation, recombination, and electrical field dependency of carrier transport characteristics is desired. Implementation of the solution of semiconductor equations in a circuit simulator enables the direct integration of the optoelectronic device level models with the electronic circuit models. This method allows for the prediction of the overall large-signal transient response by solving all OEIC variables simultaneously and self-consistently. The availability of local photogeneration models at every grid point enables the accurate simulation of optoelectronic devices without any restriction on light absorption characteristics. This simulation technique is not device-specific and can be applied to any electronic or optoelectronic device in arbitrary OEIC configuration. Since the model requires no modifications within the internal structure of the circuit simulator it can be easily implemented in different circuit simulators and computation platforms.
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Chapter 1

Introduction

In this information age, there is an increasing need for faster data collection, transmission, and processing in both analog and digital applications. Electronic and optoelectronic semiconductor devices have the dominant roles in providing the necessary systems to satisfy this growing need. Today’s high performance optical communication and signal processing systems require the monolithic integration of ultrafast optoelectronic devices with state-of-the-art electronic components. In such Opto-electronic Integrated Circuits (OEIC) not only do the individual optoelectronic and electronic devices need to be optimized, but also their circuit integration has to be considered to achieve the highest possible performance.

Actual device and integrated circuit fabrication is expensive and time-consuming. The improvement in performance comes with the increased complexity of novel devices. For such structures, detailed investigation and analysis of complex phenomena that can not be achieved through experiments nor through simple analytic models can be sought by device-level simulations. Such approaches also provide valuable insight into the physical mechanisms which ultimately determine the operational performance of the device. Most of the existing device simulators are based on the drift-diffusion (DD) description of carrier transport and employ specialized techniques to obtain the one-, two- or three-dimensional transient and steady state solutions of these equations.

The simulation of optoelectronic devices however presents some specific challenges that are not well addressed by the existing device simulation tools. One important requirement is that the local absorption and photo-generation (light-induced carrier generation) effects should be explicitly representable at any point within the device. The simulation tool should enable simple modeling of optoelectronic devices which may have localized optical absorption in specific regions and should be capable of fast transient response simulation. Besides, the tools should allow simple integration of circuit-level and device-level simulation tasks especially for the analysis of integrated circuit response under various conditions. Finally, simple extraction of device and circuit model parameters is desirable to aid the development of compact
simulation models for novel optoelectronic devices.

There exists a growing consensus that process/device and circuit simulation tools used as separate entities need to be integrated into a combined simulation environment for efficient design and optimization of semiconductor devices and integrated circuits. Such integration will shorten the development cycle for new technologies by providing accurate predictions of various physical phenomena on the device and circuit level and by allowing the investigation of detailed device behavior within the circuit environment. The majority of the proposed tool integration approaches seek either to create specific mixed-mode simulation environments or to combine specialized device and circuit simulation tools through user-friendly interfacing environments.

In this thesis a novel approach is presented for incorporating the transient solution of one-dimensional semiconductor drift-diffusion equations within a general circuit simulation tool eliminating the need for interfacing various device and circuit simulation environments. This approach allows the simple representation of localized carrier transport models of simulated devices using equivalent circuit elements such as voltage-controlled current sources and capacitors. The availability of local photogeneration models at every grid point enables accurate transient simulation of various optoelectronic devices without any restrictions on light absorption and carrier generation characteristics. The approach is conceptually simple and easy to implement in many existing circuit simulation environments with only a moderate amount of additional effort. Besides the inherent advantage of providing mixed-mode device and circuit simulation capability embedded in a circuit simulator this approach can also be used for extracting the device- and circuit-level model parameters.

In the following chapter background information about conventional mixed-mode simulation techniques and earlier work on circuit representation of semiconductor devices is given. Chapters 3 and 4 present the discretization of the time-dependent semiconductor equations a detailed derivation of model parameters and the implementation of the drift-diffusion model using equivalent circuit elements.

The application of the new approach is examined in chapter 5 with several device and mixed mode simulation examples. First a demonstration of carrier transport models is presented by analysis of absorption of an optical pulse and consequent drift and diffusion of holes in an n-type GaAs slab. The response of a simple GaAs p-i-n photodiode is examined under optical pulse illumination. Transient variation of the electron and hole concentrations, electrostatic potential, mobility and photodiode current are displayed. Then the model is used to verify that a resonant cavity enhanced (RCE) heterojunction p-i-n photodiode outperforms its conventional counterpart. A 35% bandwidth improvement and a nearly three-fold enhancement in the bandwidth-efficiency product was predicted.

Two mixed-mode OEIC simulation results are presented for our model. An Al-GaAs/GaAs heterojunction p-i-n biased by a simple RC circuit and an integrated photoreceiver. The heterojunction p-i-n is simulated for high optical excitation levels which consequently alter the photodiode bias. Hence large signal capability is
demonstrated for the model. The photoreceiver circuit consists of a simple GaAs p-i-n photodiode and an amplifier circuit utilizing a bipolar junction transistor. Transient variations in the BJT collector voltage and base current as well as photodiode current and bias voltage for a large signal optical pulse are presented. The amplifier circuit illustrates switching from saturation to cutoff when illuminated by this optical pulse.
Chapter 2

Background

Optoelectronic Integrated Circuits consist of both optoelectronic devices and electronic circuits. Circuit level models to represent the transient behavior of electronic devices have been developed. Most optoelectronic devices however lack circuit models to describe their transient characteristics, especially under large signal conditions. Therefore, to investigate the accurate behavior of devices operating within a circuit environment, mixed-mode device/circuit models are necessary. For example, in case of a photoreceiver, an electronic circuit is integrated with a photodetector. When the photodetector is illuminated by an optical pulse, the photocurrent response is detected and processed by the circuit's electronics (Fig. 2.1). In addition to processing the signal, the electronic circuit is also responsible for electrical biasing of the optoelectronic component. In the case of large optical signals, a transient variation can be observed in the bias conditions for both the electronic and optoelectronic components. Hence, an accurate mixed mode device/circuit simulation technique must reflect the device response considering changing bias conditions.

2.1 Conventional mixed-mode simulation

Many different mixed-mode simulation techniques have been demonstrated [13, 16]. In these conventional approaches, two separate simulation tools which are both installed on one platform and running simultaneously are utilized (Fig. 2.2). Each simulator is dedicated to either circuit simulation or device simulation of the mixed-mode integrated circuit. The independent circuit and device simulators are interfaced via a third simulation tool devoted to ensure the self-consistency of the mixed-mode simulation.

For a given set of bias conditions, each of the circuit and device simulation tools solves for its variables locally. Once a steady state solution is obtained, the common voltage and current variables are compared by the tool interface. Local variables are repeatedly solved until the common variables arrive at a consistent solution for both of the dedicated simulation tools. Thus, time consumption can be a serious setback.
for conventional mixed-mode simulations.

Since large signal input values alter the bias conditions for an optoelectronic device, it is nontrivial for such a mixed-mode method to converge to a result. This would require extensive communication between the separate device simulator and circuit simulator to agree on a self-consistent solution.

2.2 Circuit representation of devices

A solution to the mixed-mode device/circuit simulation problem would be to integrate the device simulation in a circuit simulating environment. Circuit simulators have been investigated since the early 1970’s, and have become nearly perfected user-friendly environments widely used by many researchers.

An example of earlier work on representing semiconductor devices in terms of lumped circuit elements is known as the Circuit Technique for Semiconductor Analysis.
(CTSA) by Sah et al. [14Γ2]. The CTSA method featured analytical as well as numerical solutions to the five main semiconductor equations—the electron and hole continuity, electron and hole current, and Poisson’s equations. Also a sixth, the trapping-charge kinetic equation [14Γ] was incorporated. Boundary conditions were applied only at points representing the external contacts. Spatial variation of impurity doping profile and the dependence of mobility on dopant impurity concentration were included. The recombination terms were modeled by multiple-energy-level Shockley-Read-Hall recombination centers.

In the CTSA model all physical phenomena are represented by conventional circuit elements via an exact transmission-line circuit model (Fig. 2.3). The circuit shown in this figure represents the steady state behavior of a one-dimensional device. The two series branches in this configuration represent the electron and hole conduc-
Figure 2.3: An example of classical transmission-line representation of the semiconductor equations from Ref. [2]. Note the non-linear bridging capacitors at every grid node complicating the simulation.

tion currents. The center node in each lumped element is the electrostatic potential for the grid point. The resistors and capacitances shown in this structure are not only nonlinear circuit elements but have values that vary with both time and space as well.

For the small signal circuit representation of the same approach the complexity increases even more. In this case an additional voltage controlled current source proportional to the generation-recombination rate is connected in parallel with each resistor shown in Fig 2.3. Implementation of this complex structure with bridging capacitors between each node in a circuit simulator would be a very challenging task.

In the model described in this thesis we demonstrate a simple circuit representation of semiconductor optoelectronic devices in which the bridging capacitors and the nonlinear circuit elements between adjacent grid nodes are eliminated. Thus the complexity of the circuit decreases allowing the model in our approach to be incorporated in a circuit simulating environment.
Chapter 3
Discretization of Semiconductor Equations

Carrier transport in semiconductors can be described by three coupled partial differential equations, namely Poisson’s equation, the electron-current continuity equation and the hole-current continuity equation. Assuming that the carrier flow and the potential variation within a semiconductor structure are confined to one dimension ($x$) the normalized carrier transport equations are given as follows.

\[
\frac{\partial^2 V}{\partial x^2} \cdot \frac{1}{r} \left[ n - p + N_A - N_D \right] = 0 \tag{3.1}
\]

\[
\frac{\partial n}{\partial t} = \frac{1}{q} \frac{\partial J_n}{\partial x} - R_n + G_n \tag{3.2}
\]

\[
\frac{\partial p}{\partial t} = -\frac{1}{q} \frac{\partial J_p}{\partial x} - R_p + G_p \tag{3.3}
\]

The normalized electron and hole current densities $J_n$ and $J_p$ are defined as

\[
J_n = \mu_n \frac{\partial n}{\partial x} - n \mu_n \frac{\partial (V + V_n)}{\partial x} \tag{3.4}
\]

\[
J_p = -\mu_p \frac{\partial p}{\partial x} - p \mu_p \frac{\partial (V - V_p)}{\partial x} \tag{3.5}
\]

Here, the normalized variables $V$, $n$, and $p$ represent the local electrostatic potential, the electron concentration, and the hole concentration, respectively. The parameter $r$ represents the normalized dielectric constant $\varepsilon R$ and $G$ represent carrier recombination and generation rates, respectively. Two potentials $V_n$ and $V_p$ are also introduced as band parameters to account for the bandgap variations in heterostructures, and are defined as follows.

\[
qV_n = \chi - \chi_r + kT \ln \left( \frac{N_C}{N_{Cr}} \right) \tag{3.6}
\]
where $N_C$ and $N_V$ are the effective density of states for conduction and valence bands respectively, $\Gamma$ is the electron affinity, $E_G$ is the bandgap and subscript $r$ denotes the reference material, e.g., GaAs or AlGaAs.

To obtain the solution of the semiconductor equations (3.1–3.3) by using the finite difference method, the equations need to be discretized on a non-uniform one-dimensional grid (Fig. 3.1). Usually, the discretization of the time-dependent semiconductor equations is carried out both in time-domain and in space-domain [17]. In the following, the dependent variables will be discretized with respect to the space (position) variable $x$, but they will be left continuous in time resulting in a semi-discrete form. The goal of this formulation is to facilitate a simple implementation of the device model within the circuit simulation environment. Three new local variables are defined as $x_1(j) = V(j), x_2(j) = V(j) - \psi_n(j)$, and $x_3(j) = -V(j) + \psi_p(j)$ where the normalized variables $V_n$, $\psi_n$, and $\psi_p$ represent the local electrostatic potential, the electron and hole quasi-Fermi potentials respectively. The Scharfetter-Gummel approximation has been used to express the current densities $J_n$ and $J_p$ in terms of the carrier concentrations [9]:

$$J_n(j + \frac{1}{2}) = \frac{\mu_n(j + \frac{1}{2}) \Delta_n(j)}{h_{j+1}} \left[ \frac{n(j+1) \exp(\Delta_n(j)) - n(j)}{\exp(\Delta_n(j)) - 1} \right] \quad (3.8)$$

$$J_p(j + \frac{1}{2}) = \frac{\mu_p(j + \frac{1}{2}) \Delta_p(j)}{h_{j+1}} \left[ \frac{p(j+1) \exp(\Delta_p(j)) - p(j)}{\exp(\Delta_p(j)) - 1} \right] \quad (3.9)$$

where

$$\Delta_n(j) = V(j) + V_n(j) - V(j+1) - V_n(j+1) \quad (3.10)$$

$$\Delta_p(j) = V(j+1) - V_p(j+1) - V(j) - V_p(j) \quad (3.11)$$

The grid spacing at the $j$th grid is defined as $h_j$. The carrier concentrations $n$ and $p$ are related to the local potential variables as

$$n(j) = \exp[V(j) + V_n(j) - \psi_n(j)]$$

$$= \exp[x_2(j) + V_n(j)] \quad (3.12)$$

$$p(j) = \exp[-V(j) + V_p(j) + \psi_p(j)]$$

$$= \exp[x_3(j) + V_p(j)] \quad (3.13)$$

Substituting (3.8–3.13) into the device equations (3.1–3.3), a set of time-dependent space-discretized equations are obtained on a nonuniform one-dimensional grid (Fig. 3.1).
Figure 3.1: Partitioning of a $p - n - n^+$ structure on a one-dimensional non-uniform grid.

\[
x_1(j) = \frac{h_j h_{j+1}}{r_{j+\frac{1}{2}} h_j + r_{j-\frac{1}{2}} h_{j+1}} \frac{h_j + h_{j+1}}{2} \left[ \exp(x_3(j) + V_p(j)) - \exp(x_2(j) + V_n(j)) + C_j \right] \\
+ \frac{r_{j+\frac{1}{2}} h_j x_1(j+1) + r_{j-\frac{1}{2}} h_{j+1} x_1(j-1)}{r_{j+\frac{1}{2}} h_j + r_{j-\frac{1}{2}} h_{j+1}}
\]

(3.14)

\[
\frac{\partial x_2(j)}{\partial t} = \frac{2 \mu_{n(j+\frac{1}{2})}}{(h_j + h_{j+1}) h_{j+1}} \left[ x_1(j) - x_1(j+1) + V_n(j) - V_n(j+1) \right] \\
\times \left\{ \frac{\exp[x_2(j+1) - x_2(j)] + x_1(j) - x_1(j+1)}{\exp[x_1(j) - x_1(j+1) + V_n(j) - V_n(j+1)] - 1} \right\}
\]
and Eqns. (3.15) correspond to the electron continuity and hole continuity equations, respectively. The introduction of band gap parameters $V_n$ and $V_p$ allows simple representation of the band gap variation effects in heterostructures [9]. Notice that a photon-induced carrier generation term $G_j$ is available at every grid point $x_j$, and it can be represented by an independent current source. The extension of this approach for two-dimensional semiconductor equations is also straightforward. Assuming a rectangular space-grid structure, the Poisson’s equation, the electron continuity equation, and the hole continuity equation can easily be discretized at each grid point. In this case, each discretized equation contains the electrostatic potential and Fermi potential variables of the four (instead of two) neighboring grid points. Thus the resulting set of time-continuous and space-discretized equations will again be characterized by next-neighbor relations in the space domain.

### 3.1 Recombination models

The carrier generation/recombination rate at any grid point is assumed to be a function of two basic mechanisms, i.e., trap-related phonon transitions and photon transitions. The four partial processes involved in phonon transitions are: (i) electron capture from the conduction band by an unoccupied trap, (ii) hole capture, (iii) release of an electron from an occupied trap to the valence band, and (iv) electron emission from an occupied trap to the conduction band.
from the valence band by an unoccupied trap. The well-known Shockley-Read-Hall model accounts for the total generation/recombination rate due to phonon transitions listed above [12].

\[ R_j = \frac{n(j)p(j) - n_i^2}{\tau_p(n(j) + n_i) + \tau_n(p(j) + p_i)} \]  

(3.17)

where \( n(j) \) and \( p(j) \) represent the electron and hole concentrations at the jth grid point and \( \tau_n \) and \( \tau_p \) represent the carrier lifetimes respectively.

The recombination term \( R_j \Gamma \) can also be expressed in terms of electrostatic potential and the Fermi potentials at the jth grid point as follows [17] \( \Gamma \)

\[ R_j = \frac{\exp(x_2(j) + V_n(j))\exp(x_3(j) + V_p(j)) - 1}{\tau_p(\exp(x_2(j) + V_n(j)) + n_i) + \tau_n(\exp(x_3(j) + V_p(j)) + p_i)} \]  

(3.18)

In the model, Eqn. 3.18 has been utilized to describe recombination in semiconductor devices. Both of the electron and hole lifetimes were assumed to be 1 ns.

The photon transitions on the other hand are assumed to be exclusively in the form of optical generation of carriers described as an electron gaining energy from incident photons and moving from the valence band to the conduction band. This process is represented by the photon-induced carrier generation term \( G_j \) at each grid point.

### 3.2 Mobility models

In Equations (3.14–3.16) the variation of electron and hole mobilities as a function of doping concentration, alloy composition, and local electric field is taken into account by a suitable local mobility model based on carrier velocity saturation. Approximations of these effects derived for the GaAs material systems [9] suitable for numerical simulation is utilized in our device model.

In order to calculate the hole mobility for an arbitrary doping and composition, an empirical formula \( \mu_{p-GaAs} \) is utilized to approximate the hole mobility of GaAs. The form of \( \mu_{p-GaAs} \) is

\[ \mu_{p-GaAs}(T, N_D + N_A, E_{Qp}) = \frac{\mu_{p-L}}{1 + 3.17 \times 10^{-17}(N_D + N_A)[0.206]} \times \left\{ \frac{300}{(1 + E_{Qp} / [(1.95 \times 10^4)T]^2.7]} \right\}^{\frac{cm^2}{V \cdot sec}} \]  

(3.19)

where \( E_{Qp} \) is the effective field for holes readily calculated by the model \( \mu_{p-L} \) is the low field and intrinsic mobility for the material while \( T \) denotes the temperature. In the above formula \( \mu \) the field and temperature dependent portions are assumed to be the same as those for p-type silicon.
The absence of a well structured experimental characterization of the electron mobility as a function of composition, carrier compensation, doping and applied field make the development of an accurate model difficult. An electron mobility model developed in Ref. [9] has explicitly considered only electrons which occupy the $\Gamma$ and the X-valleys. The electron mobility in GaAs is taken as

$$\mu_{n-GaAs}(T, N_D + N_A, E_{Qn}) = \frac{\mu_{LD} + v_{sat} \left| E_{Qn} \right|^3 / E_{C-GaAs}^4}{1 + \left( \left| E_{Qn} \right| / E_{C-GaAs} \right)^4}$$

(3.20)

where the low field doping concentration dependent mobility is

$$\mu_{LD} = \frac{\mu_{n-L}}{\left[ 1 + 5.51 \times 10^{-17} (N_D + N_A)^{0.233} \left( \frac{300}{T} \right)^{2.3} \right]}$$

(3.21)

cm$^2$/V-sec,

the saturation velocity is

$$v_{sat} = (1.28 - 0.0015 T) \times 10^7 \frac{cm}{sec},$$

(3.22)

and the critical field is

$$E_{C-GaAs} = (5.4 - T/215) \frac{kV}{cm}.$$

(3.23)

In our model GaAs is used as the reference material system. Thus all InGaAs, AlGaAs materials simulated are assumed to have the same electric field dependency as GaAs. For small mole fractions of Al and In this is a valid approximation. The intrinsic/low field mobilities $\mu_{LD}$ and $\mu_{p-L} \Gamma$ are for each grid to characterize the material used.

In Fig. 3.2 the dynamic behavior of the electron mobility due to the variation in the electrostatic potential is illustrated for a photodiode under large signal optical excitation. The mobility variations of three different grid points in a GaAs p-i-n photodiode are shown. Figure 3.2 indicates the importance of incorporating the field dependence especially for investigating the transient effects of semiconductor devices.
Figure 3.2: Transient behavior of the electron mobility at three different grid points due to variation of the electrical field in a photodiode.
Chapter 4

Implementation of the Model

The semi-discrete carrier transport equations (3.14–3.16) describe the transient behavior of the electrostatic potential and the Fermi potentials at any grid point. These equations have the following general form:

\[
x_1(j) = f_1[x_1(j+1), x_1(j-1), x_2(j), x_3(j), V_n(j), V_p(j)]
\]

\[
\frac{\partial x_2(j)}{\partial t} = f_2[x_1(j), x_1(j+1), x_1(j-1), x_2(j), x_2(j+1), x_2(j-1),
\]
\[
x_3(j), G(j), V_p(j), V_n(j), V_n(j-1), V_n(j+1)]
\]

\[
\frac{\partial x_3(j)}{\partial t} = f_3[x_1(j), x_1(j+1), x_1(j-1), x_3(j), x_3(j+1), x_3(j-1),
\]
\[
x_2(j), G(j), V_n(j), V_p(j), V_p(j-1), V_p(j+1)]
\]

It is seen that the electrostatic and Fermi potentials at any grid point are described by three coupled nonlinear differential equations in the time domain which involve the corresponding potential variables as well as the neighboring grid node potentials. With a simple variable assignment the equations (4.1–4.3) can be represented by an equivalent circuit which can then be solved using a conventional transient circuit simulation program such as SPICE. For a simple equivalent circuit interpretation of these equations first let the variables \(x_1\), \(x_2\), and \(x_3\) represent three node voltages. Also let the right-hand sides of equations (4.1–4.3) the functions \(f_1\), \(f_2\) and \(f_3\) represent the currents of three nonlinear voltage-controlled current sources. Then equations (4.2) and (4.3) can be interpreted as the state equations of linear capacitors with terminal currents of \(f_2\) and \(f_3\) respectively while equation (4.1) describes the terminal voltage of a unit resistor with a terminal current of \(f_1\). The equivalent circuit diagrams corresponding to equations (3.14–3.16) are given in Fig. 4.1. It is seen that each space-grid point is represented by three circuit nodes where the node potentials correspond to the electrostatic potential and the two Fermi potentials associated with
Figure 4.1: The one-dimensional finite difference grid and the equivalent-circuit representation of the variables associated with each grid point. All resistors and capacitors shown in each grid are taken as unity.
the grid point.

The conversion of the semiconductor drift-diffusion equations into an equivalent circuit representation was first proposed by Sah [14] as described in chapter 2, and later pursued by others [24]. In the modeling approach presented here in contrast the local circuit nodes representing grid points are linked through node voltages only. The new formulation approach also eliminates the bridging capacitors and the nonlinear circuit elements used in previous circuit element representations. The modularity of the new modeling approach can also be utilized for adjusting the grid point density of the simulated structure [7]. The boundary conditions are established assuming purely ohmic contacts at both external boundaries. The electrostatic potential at the boundary is determined by the externally applied bias voltage which in turn depends on the external load circuit. It is assumed that no voltage drop occurs at the boundary. The conditions for the carrier concentrations at the ohmic boundaries are established by increasing the surface recombination rate. This is accomplished by gradually reducing the electron and hole lifetimes in the Shockley-Read-Hall recombination term by three orders of magnitude for the grid points in the immediate neighborhood of the boundaries.

The calculation of currents is also carried out using the equivalent circuit representation. The total conduction current component across the device is found as the sum of electron and hole currents \( J_n \) and \( J_p \).

\[
J_c = J_n(x_j) + J_p(x_j)
\] (4.4)

The displacement current component is calculated as

\[
J_d = -\frac{1}{A} C_{jnct} \frac{d}{dt} V_{space-charge}
\] (4.5)

where \( C_{jnct} \) represents the junction capacitance and the space-charge voltage is found by integrating the local carrier densities. The displacement current component for given grid node can be computed directly by taking the time derivative of the local electric field.

\[
J_d = -\epsilon \frac{d}{dt} E
\] (4.6)

where \( \epsilon \) is the dielectric constant. In the presented model, the electrical field is obtained by the difference of the electrostatic potential in the neighboring nodes and time derivation is implemented by a simple circuit involving a voltage source and a capacitor. At every time step, therefore, we readily have all the current components and the total current.

Alternatively, the displacement current can also be evaluated by integrating the time derivative of carrier concentrations in the depletion region.

\[
J_d = \frac{C_{jnct} q}{\epsilon A} \int_0^{x_m} dy \int_0^y \left[ \frac{\partial p(x, t)}{\partial t} - \frac{\partial n(x, t)}{\partial t} \right] dx
\] (4.7)
This formulation is particularly useful when the transient variation of electrostatic potential is difficult to evaluate. In the present model we utilize direct calculation of the drift current at given grid node using (4.6).

Given the device structure, grid spacing and the relevant model parameters, the equivalent circuit representation of the semiconductor drift-diffusion equations can thus be readily generated. A conventional circuit simulation program is then used to solve these equations in the time domain and to obtain the variation of the electrostatic potential and the Fermi potentials at each grid point. The approach proposed here also makes full use of the numerical stability and efficiency of the well-established circuit simulation routines.

The exponential terms which appear in the space-discretized Poisson's equation (3.14) and the electron and hole continuity equations (3.15–3.16) essentially possess the same characteristics as the compact diode model or the BJT Ebers-Moll model equations which are handled quite well numerically in most conventional circuit simulators. The Bernoulli equation appearing in Eqs. (3.15) and (3.16) which has the general form

$$f(x) = \frac{x}{\exp(x) - 1}$$  \hspace{1cm} (4.8)

presents a challenge in numerical evaluation for very small values of $x$ which is encountered in neutral regions because the numerator and the denominator approach zero in this case. However, since the function does not possess a singularity at $x = 0$ the problem can easily be circumvented by assigning the function its known limit value $f(0) = 1$ for $x$ values smaller than a certain lower boundary [7]. The availability of the absolute value function for non-linear controlled sources in SPICE provides the alternative option in which (4.8) is modified as

$$f(x) = \frac{|x| + \delta}{|\exp(x) - 1| + \delta}$$  \hspace{1cm} (4.9)

where $\delta$ is a small positive number. The value of $\delta$ (typically $10^{-6} - 10^{-8}$) can be arbitrarily set to control the precision of the equations. A more detailed numerical error analysis is not presented here mainly because the error analysis results derived for conventional circuit simulation tools can be applied directly in this case.

The simple equivalent circuit representation allows detailed modeling of the internal device structure by locally specifying the physical device parameters. This model allows the investigation of the transient behavior of optoelectronic devices using the local photo-generation term $G_j$ [14Γ20Γ21].

It was already mentioned that mixed device/circuit simulation capability using the same simulation tool is very desirable for accurate and realistic assessment of novel devices and integrated circuits. Since the external voltage and current conditions of the simulated device are available at any time point $t$ and since the discretized semiconductor equations are being solved using circuit simulation routines, the device
simulation approach presented here lends itself to mixed-mode simulation of devices and circuits by directly linking device-level and circuit-level models. The device simulation model described above has been implemented and tested successfully in SPICE and two SPICE-like general-purpose circuit simulation programs. Figure 4.2 shows the major components of the combined device/circuit simulation environment.

To facilitate initial DC convergence during the simulation of some semiconductor structures, a simple ramping scheme is used. During preliminary studies the entire structure is initially defined as having the same uniform doping density with zero potential drop across its external terminals. The DC convergence for this uniform structure can be easily achieved since the local electrostatic potential and the electron and hole densities will be constant and equal at all grid nodes. Next, the local doping densities and the external boundary conditions are gradually changed (ramped) in the time domain until they reach their actual levels intended for the semiconductor structure. At this point the system is in steady-state and the actual excitation can now be applied either as an electrical or optical pulse waveform.

For the implementation in SPICE, DC convergence for a p-n junction can be achieved without any pre-conditioning or ramping for moderate doping densities and applied bias. Although there is no formula for the maximum allowable values and they are usually interdependent, typical numbers are in the order of $10^{17} \text{cm}^{-3}$ and $50kT$ for doping and applied bias respectively for a p-n junction. In case of failure to achieve DC convergence, ramping of one or more device parameters can be utilized.

It was also found that abrupt changes in electrical or optical excitation waveforms occasionally cause numerical convergence problems during simulation. To avoid such problems a “precursor” pulse several orders of magnitude smaller than the actual excitation pulse may be applied first. This small excitation ensures that the carrier density levels rise sufficiently above the steady-state levels so that the subsequent excitation pulse does not cause numerical problems. Since the magnitude of the precursor pulse is very small it does not influence the transient response of the device in any significant way. We also noted in SPICE that if the rising edge of the optical pulse has a staircase structure of arbitrarily small time step size convergence is assured. In this scheme pulses with very small rise time (sub-picosecond) can be accurately represented.

## 4.1 Obtaining device descriptions

In order to achieve the circuit representation of a given device seven numerical values of the device characteristics must be given. Once the number of grids the device is partitioned by is specified for each grid; 1) grid spacing 2) dielectric constant 3) the intrinsic/low field electron mobility 4) the intrinsic/low field hole mobility 5) conduction band discontinuity 6) valence band discontinuity 7) intrinsic concentration must be specified. These values for the simulations described in the next chapter
Figure 4.2: The proposed mixed-mode device/circuit simulation environment. No additional software tools are required for integration of device and circuit simulation environments.
were obtained from Ref. [10].

For a fifty node device inserting all of the 350 variables manually would be neither time efficient nor reliable. Thus, a C program has been developed that reads the device description from a data file, computes all of the coefficients of the device models, and creates a SPICE-ready circuit file. Therefore, the preprocessor described in fig. 4.2 which changes the device description to circuit representation is accomplished by this C program. This program has been optimized to create as few nodes in the circuit file as possible and set boundary and initial conditions. Therefore the computation load is minimized and the convergence of SPICE is facilitated. The interface allows the device bias voltage, generation grid numbers, transient analysis specifications and tolerances to be imported into the circuit file.
Chapter 5

Simulation Examples

The physical phenomena incorporated in the proposed device simulation approach will be demonstrated in this chapter by simulation examples. Demonstration of this model was initially carried out in SPICE-like general-purpose circuit simulation programs [10]. Recently the model was also implemented in SPICE [20-21] and we have noted improved capabilities. The results presented are obtained using SPICE version 3F2. Though the simulation results presented in this chapter prove the model to be qualitatively correct, one may wish to do an analysis on its results quantitatively.

5.1 Carrier transport models

The first example investigates the drift-diffusion and recombination of excess carriers generated by a localized light pulse in an n-type semiconductor sample. The semiconductor sample has a length of $L = 2\mu m$ and a doping concentration of $N_D = 10^{16}$ cm$^{-3}$. A voltage of $50 kT/q$ is applied across the sample and a light pulse focussed on the mid-point of the sample is used to generate excess carriers (Fig. 5.1(a)). Note that this arrangement corresponds to the well-known Haynes-Shockley experiment for the measurement of carrier drift mobility in semiconductors. Figure 5.1(b) shows the excess hole concentration within the sample as a function of position and time. It is observed that the generated excess carriers diffuse away from the mid-point and recombine while the concentration peak moves (drifts) toward the negative-biased end of the sample under applied electric field. This example qualitatively demonstrates that the equivalent circuit models introduced in chapter 3 represent the transient drift-diffusion and recombination behavior of carriers in a semiconductor sample.
Figure 5.1: Uniform n-type semiconductor sample being subjected to a localized optical pulse excitation leading to excess carrier generation in a narrow region (top). Simulated variation of hole concentration as a function of position and time. Generated excess carriers diffuse away from the narrow generation region while the concentration peak drifts toward the negative-biased end of the sample under applied electric field (bottom).
5.2 GaAs p-i-n photodiode

This example investigates the drift and diffusion of excess carriers generated by a localized light pulse in a GaAs p-i-n diode structure (Fig. 5.2). The total thickness of the device is $L = 1.5\mu m$ and low doped normally depleted region is $1\mu m$ thick. A total of 50 grid points were taken with variable spacing i.e. finer grids at the metalurgical junctions. A small voltage of $10kT/q$ is applied across the sample to reverse-bias the diode and a light pulse with 0.1 ps rise- and fall-times and 50 ps duration is focussed on the depletion region of the diode.

For the demonstration of the concept uniform optical generation is assumed to take place in the $0.5\mu m$ thick region located in the middle of the depletion region. Assuming a 40% quantum efficiency (typical for $0.5\mu m$ GaAs absorber) the optical excitation corresponds to $5 \times 10^{21}$ photons/cm$^2$-s (i.e. $1kW/cm^2$). Although the power density seems to be extremely large the corresponding total power is in the order of 10mW for a $30 \times 30\mu m$ device and for the given pulse duration total energy per pulse is only 50fJ.

Figures 5.3 and 5.4 show the variation of electron and hole concentrations...
Figure 5.3: Simulated variation of electron concentration in a GaAs p-i-n photodiode as a function of time and position.

respectively within the sample as a function of position and time. As can be seen from the carrier concentration plots, the optical generation is large enough to create electron-hole densities comparable to those in the neutral regions demonstrating the large signal capability of the presented model. A comparison of transient variation of electron and hole concentrations reveals that electrons, due to higher mobility, traverse the depletion region much faster than the holes.

In the model, field dependence of the material parameters were taken into account as dynamic variables as mentioned in section 3.2. To demonstrate the importance of transient consideration of the discussed material properties, such as mobility, Figures 5.5 and 5.6 show the time and space dependence of the electrostatic potential and electric field respectively under the same optical excitation for the photodiode. For the given excitation level, the variation of electric field during the pulse is very prominent. Therefore, the corresponding variation in carrier dynamics especially in
Figure 5.4: Simulated variation of hole concentration in a GaAs p-i-n photodiode as a function of time and position.

The electron mobility is very significant (see Fig. 3.2).

The terminal current for the photodiode is calculated at every time step using dedicated circuits inside the equivalent device model. The conduction and displacement current components for the photocurrent response of the GaAs p-i-n photodiode described above is shown in Fig. 5.7. For this simulation run a 20 ps 1 μW square optical pulse was absorbed at the n-side of the depletion region. The delay between the optical pulse and the conduction current is due to long transit time for holes. Besides, a significant fraction of the holes generated in the proximity of the depletion edge diffuse into the neutral n-region as shown in Fig. 5.4. For an absorption profile close to n-region the response speed is further limited by the slow diffusion process.

For two different absorption profiles the photocurrent responses of the GaAs p-i-n is shown in Fig. 5.8. The photocurrent with faster response time corresponds to
Figure 5.5: Simulated variation of electrostatic potential in a GaAs p-i-n photodiode as a function of time and position.

absorption closer to the p-contact ($\sim 0.15\mu m$ from the depletion edge) resulting in a shorter distance to transit for holes than that for electrons. In this case the rise and fall times of the current are 8 ps and 9 ps respectively. Therefore a photodetector with such an absorption profile will have a bandwidth approaching 50 GHz. In contrast for absorption at the opposite end of the depletion region ($\sim 0.15\mu m$ from n-region) the photocurrent response has a fall time of 40 ps due to the longer transit time of holes. This corresponds to a bandwidth of 9 GHz.

5.3 Resonant cavity enhanced photodetectors

As described earlier the availability of photogeneration terms at every grid point in our model enables the study of novel photodetector structures with arbitrary light absorption profiles. In this section the resonant cavity enhancement (RCE) effect
and its implications on high speed devices are presented. The implementation of RCE scheme in a heterojunction p-i-n photodiode is described. The superiority of the RCE detection scheme is confirmed by comparing its performance to a conventional p-i-n photodiode via the simulation tool described in this thesis.

A large quantum efficiency $\eta$, i.e., the probability of detecting incident photons, is an important property for high performance photodetectors. The quantum efficiency of conventional detector structures is governed by the absorption coefficient of the semiconductor material requiring thick active regions for high quantum efficiencies. However, reduced device speeds result from the long transit times required in devices with thick active regions. It is desirable to enhance the quantum efficiency without increasing the active layer thickness in order to optimize the gain-bandwidth product.

In the RCE photodetectors, the detection region is integrated into an asymmetric Fabry-Perot cavity formed by two mirrors (top and bottom) and at the resonance
condition the incoming light interferes constructively with the reflected component from the bottom mirror. The resulting resonant cavity effect enhances the internal optical field amplitudes at the resonant wavelengths $[18\Gamma6]$. Resonant cavity enhancement can also be viewed as a multiple-pass detection over a thin absorbing region. For a high quality factor $Q$ cavity (high mirror reflectivities and a thin absorption region) a drastic increase in $\eta$ can be obtained.

For comparison of the conventional and RCE detectors we have selected two heterojunction $p$-$i$-$n$ photodiode structures. These structures follow the design rules suggested in Ref. $[22]$ for highest performances. The intrinsic region is assumed to be lightly $n$-doped and its width is selected to be $0.72\mu$m corresponding to the maximum bandwidth-efficiency product condition for a $10 \times 10\mu$m$^2$ conventional $p$-$i$-$n$ diode. This depletion width is slightly larger than the optimum value for the RCE-detector. Therefore this selection favors the conventional structure.

Figure 5.7: Detail of the photocurrent displaying transient behavior of displacement (dotted-line) and conduction (dotted-dashed line) components. Dashed line shows the generation term.
The conventional device which we modeled (#1) has a 0.64 μm thick normally depleted n^-GaAs absorbing region and Al_{0.06}Ga_{0.94}As contact (or window) layers. The schematic diagram of the device is shown in Fig. 5.9(a). The AlGaAs contact layers allow for high speed operation by removing the limitation imposed by the current diffusion out of absorbing but undepleted regions which is a slow process. Within a wavelength range (825nm < λ ≤ 870nm) determined by the absorption edge of the AlGaAs and GaAs layers, photogeneration will occur only in GaAs regions (α = 10^4 cm^-1). The carriers that may be generated in the GaAs substrate and cap layers are blocked by the AlGaAs barriers i.e., they do not contribute to the photocurrent and thus the device speed is limited solely by the transit time of photogenerated carriers across the 0.72 μm depletion region. A small Al mole fraction was chosen to avoid large band discontinuities. The heterojunctions were graded and positioned
inside the depletion region to prevent charge trapping. The total band discontinuity is 75 meV/2/3 of which is assumed to be at the valence band. For this conventional detector structure we assume a nearly ideal anti-reflection coating (surface reflectivity ≈ 0.05) resulting in a quantum efficiency of 0.45 within the 825-875 nm wavelength range.

The RCE photodiode structure (Device# 2) has GaAs contacts (p and n) and depletion regions (Fig. 5.9(b)). A 0.08μm thick In0.07Ga0.93As absorption region is placed in the depletion region extending the wavelength sensitivity spectrum of the detector to 920 nm. Within a wavelength range of 870 nm to 920 nm only this InGaAs region absorbs (α = 10⁴ cm⁻¹) the incident light and the remainder of the detector is transparent.

The position of the absorbing layer is optimized so that the carriers have to traverse the depletion region proportional to their velocities. Therefore holes which have the slower carrier mobility pass through a short depletion region and electrons through a long depletion region. This ensures the arrival of the photogenerated carriers (at the absorption region) at both contacts simultaneously eliminating ‘tails’ in the current responses which limit the bandwidth of conventional detectors. For this optimization we first run the device simulation with no optical excitation and calculate the electron and hole mobilities in the depletion region for the electrical bias to be used for transient analysis. These mobilities were used to evaluate the ratio of electron and hole velocities and thus the position of the absorption region for the fastest detector response. As a result the distance of the absorption layer from the p-region is approximately half of that from the n-region.

We consider a Fabry-Perot resonant cavity formed by an ideal bottom mirror (R₂ ≈ 1.0) and a high reflectivity (R₁ = 0.7) top mirror. The quantum efficiency of this RCE-detector is estimated at η = 0.9 at wavelengths around 900 nm. The electrical contacts are assumed to be between the multi-layer mirrors to prevent high resistances. Since the current does not flow through the mirror regions the large band discontinuities required for the mirror formation does not affect the electrical performance.

Figures 5.10 and 5.11 illustrate the time evolution of the hole p(x, t) and electron n(x, t) concentrations respectively for the InGaAs/GaAs heterojunction photodiode structure (Device# 2) described above. Steady state carrier distributions as can be observed at t = 0 clearly shows the depletion region and the location of heterojunctions. The wavelength of the optical excitation was chosen so that photogeneration would occur in the InGaAs region only. The optical excitation is a square pulse of about 9 × 10¹⁴ photons/device for a 10 × 10μm² device (≈ 180W/cm²) with negligible rise and fall times (0.01ps) and a full−width−at−half−maximum (FWHM) of about 10 ps. We assume that 90% of the incident photons are absorbed in the thin InGaAs region. Such a high quantum efficiency can be achieved using the RCE-detection scheme [18Γ6Γ19].

The transient variation of electron and hole variations (Figs. 5.10 and 5.11) reveal
Figure 5.9: The schematic flat band diagram (applied and built-in fields are not shown) of the modeled p-i-n photodiodes and the qualitative representation of the photogeneration terms. (a) Device #1: the conventional AlGaAs/GaAs p-i-n heterojunction photodiode. The photogeneration occurs in the depleted GaAs region on an exponentially decaying profile. (b) Device #2: the GaAs RCE photodiode with an InGaAs absorbing region. The generation is localized in the InGaAs region. The band discontinuities are exaggerated to clearly identify the graded regions.
Figure 5.10: Simulated variation of hole concentration in the InGaAs/GaAs p-i-n diode (Device #2) as a function of time and position when light is absorbed across the depletion region. The steady state hole distribution can be observed at $t = 0$.

The advantages of the design with a thin absorbing layer embedded in the depletion region. Both electrons and holes are swept under the electrical field in the depletion region immediately after photo-generation. Carrier diffusion into the neutral regions is negligible and the carrier concentrations return to their steady state values shortly after the termination of the optical excitation. Figure 5.12 shows the short circuit photocurrent response of the same photodiode (Device #2). The time variation of the displacement and conduction current components are illustrated in comparison with the generation term. Since there are no assumptions and approximations in the time variation of the semiconductor equations in the presented model, the calculated current is an accurate representation under large pulse excitation.

The ultimate response time of photodetectors can be determined by observing their transient current under optical pulses of varying duration. Within the conver-
Figure 5.11: Simulated variation of electron concentration in the InGaAs/GaAs p-i-n diode (Device #2) as a function of time and position when light is absorbed across the depletion region.

gence capabilities of the circuit simulator we have utilized it is possible to calculate the output current for an optical pulse of less than 1 ps FWHM. Since the estimated transit time for conventional and RCE device designs are about 5 to 10 ps we compare the transient response of the two devices under a 5 ps optical pulse. To ensure small signal condition pulse amplitude was chosen as 1% of the previous example (Figs. 5.10-12). Figure 5.13 shows the simulated short circuit currents for the conventional (dashed line) and RCE (solid line) p-i-n detectors under a 5 ps FWHM optical pulse. At first glance we observe a larger (magnitude) and sharper (less time spread) current for the RCE-detector compared to the conventional case.

A close inspection of the individual current components reveals the expected transient features and superiority of the RCE-detector. (i) The rise and fall times of the total current are comparable for the RCE design. For the conventional detector the
Figure 5.12: The short circuit current output evaluated for the simulation example shown in Figs. 5.10 and 5.11 (time axis is re-scaled for clarity). The photogeneration term and displacement (dashed line) and conduction (dotted line) current components are illustrated.

Fall time is much larger than the rise time as a result of long transit time for holes.

(ii) The RCE-detector not only has larger current under identical optical excitation owing to the enhanced quantum efficiency but also the response is faster than the conventional counterpart. Therefore the bandwidth–efficiency product is doubly improved by the RCE-detection scheme. (iii) Due to the simultaneous arrival of both kinds of carriers at the contacts the time spread of the conduction current for the RCE design is much smaller than the conventional p-i-n in which the carriers reach the contacts at different times determined by the position of the photogeneration.

For a comparison of the RCE and conventional detectors in the frequency domain we used Fast-Fourier-Transform (FFT). The simulated temporal responses depicted in Fig. 5.13 were converted into frequency domain by FFT and shown in Fig. 5.14. Fourier analysis of the optical excitation term gives a flat spectrum (within 1dB) up to
Figure 5.13: The transient short circuit photocurrent under a 5 ps FWHM optical pulse for AlGaAs/GaAs conventional p-i-n (Device #1 dashed line) and GaAs/InGaAs RCE detector (Device #2 solid line).

approximately 100 GHz verifying the accuracy of the frequency domain representation of the current pulses. Furthermore, to eliminate the influence of the finite width of the optical excitation pulse, it was deconvolved from the simulated current responses. From Fig. 5.14, the transit time limited 3–dB bandwidth of the conventional p-i-n detector can be calculated as 52 GHz in comparison with 70 GHz for the RCE-pin, corresponding to a 35% improvement.

When we compare the transit time limited bandwidth–efficiency products of the two structures we see a drastic improvement. For the conventional p-i-n BWE is 23 GHz which is slightly smaller than the theoretical limit (27 GHz [22]) based on the transit time of holes. The BWE of the RCE-pin is 63 GHz which represents a nearly 3-fold improvement. If the optimum depletion width for RCE detector is considered \( L \sim 0.5 \mu m \) [22] then the BWE can be extended to 100 GHz.

For a direct comparison of the bandwidths for RCE and conventional p-i-n struc-
Figure 5.14: Frequency responses of conventional (dashed) and RCE (solid line) p-i-n photodetectors obtained by Fourier transform of pulse responses shown in Fig. 5.13. The optical excitation term was deconvolved to extract the impulse response.

We plot the peak value of the normalized short circuit current (or detector response) as a function of the inverse pulse FWHM (Fig. 6.14). When the pulse duration is large, the output current reaches its maximum value determined by the internal quantum efficiency, i.e., $\eta = 0.45$ for conventional and $\eta = 0.9$ for RCE detectors. As the optical pulse width becomes smaller, the electrical current is unable to reach the steady state value. Although the ratio of electron hole pairs to the number of incident photons remains the same, the peak current decreases since the current is spread over time. The peak short circuit current drops to half of its maximum at a pulse FWHM of 4.5 ps and 3.5 ps for conventional and RCE p-i-n detectors, respectively. Therefore, the bandwidth of the RCE p-i-n is approximately 30% larger than that of the conventional diode (see Fig. 5.15) confirming our findings through Fourier analysis. Note that both of these comparisons are made for a single optical pulse of varying duration. If we consider repetitive pulses or sinusoidal excitation, the
Figure 5.15: Detector response (peak value of the normalized detector current) versus the inverse excitation pulse width for conventional (dashed) RCE (solid line) p-i-n detectors.

bandwidth difference will be even larger due to the large fall time of the conventional p-i-n.

To this end we also compare the fall times for RCE and conventional p-i-n diodes. We calculated the response of these devices (Devices #1 and #2) under a long optical pulse such that both devices can reach the steady state current levels. The optical pulse is terminated abruptly (fall time=0.01 ps) and we observe the decay of the short circuit current for both cases as shown in Fig. 5.16. Despite the higher steady state level the current from the RCE detector decays faster and drops to lower values than the current for the conventional p-i-n. We have computed the fall time $\tau_f$ as the time it takes the current output to drop from 90% to 10% of the steady state value. The fall time $\tau_f = 7.1$ ps for the RCE p-i-n is 60% of that for the conventional structure ($\tau_f = 11.7$ ps) corresponding to 65% faster response.
Figure 5.16: The transient short circuit photocurrent under a long optical pulse with 0.01 ps fall time for conventional (dashed) and RCE (solid) p-i-n detectors. The excitations is long enough to allow both devices to reach their steady state current levels.

5.4 AlGaAs/GaAs heterojunction p-i-n

In this section an AlGaAs/GaAs heterojunction p-i-n photodetector externally biased by a simple RC circuit as shown in Fig. 5.17 is simulated. While this simulation demonstrates both mixed-mode and large signal capability the latter is emphasized in this example.

A small voltage of $V_{ext} = 0.5$ Volts is applied to the circuit which reverse biases the photodiode. The junction capacitance of the device is represented by the capacitor $C$ shown in Fig. 5.17. The total thickness of the device is $L = 1.6 \mu m$ with a low doped intrinsic region $0.8 \mu m$ thick. The highly doped GaAs cap layers representing realistic ohmic contact regions are $0.1 \mu m$ each. Utilizing variable grid spacing $50$ grid nodes were taken. Our model provides the means to observe the device variables as well as the circuit (II/V) variables during the transient simulation.
A Simulation Example:

GaAs - AlGaAs Heterojunction p-i-n Photodiode

Figure 5.17: The simulated photoreceiver circuit utilizing an AlGaAs/GaAs p-i-n heterojunction photodiode structure biased by a simple RC circuit. C=70fF and R=300 ohms.

Figure 5.18 shows the typical I-V characteristics of a photodiode under stepped illumination. The DC load line is imposed by the simple bias circuit where $V_{\text{ext}}$ and $V_{\text{ext}}/R$ are the points the load line intersects the x and y axis respectively. The bias condition on the photodetector is determined by the intercept of the load line with the diode I-V characteristics corresponding to the steady-state optical excitation. Under small signal conditions the variation in the optical signal does not have a significant effect on the bias conditions; thus the terminal voltages can be assumed constant. In contrast for large optical excitation case the I-V characteristics of the photodiode will vary significantly as shown in Fig. 5.18 resulting in changing bias conditions. Thus bias conditions can no longer be assumed as constant and they should be considered as dynamic variables. Extreme case of large signal is when optical excitation is large enough to drive the photodetector into saturation [11]. Saturation refers to the state
Figure 5.18: Photodiode I-V characteristics under increasing optical illumination

of operation where the input power is so large that the electrical output current and voltage no longer follow the input linearly. Saturation not only distorts the signal waveform but also increases the transit time as will be shown later which in turn decreases the bandwidth of the device.

For the demonstration of the concept a uniform optical generation is assumed to take place in the depletion region. The light pulse has 0.01 ps rise- and fall-times and a 20 ps duration corresponding to a $5 \times 10^{20}$ photons/cm$^2$s (i.e. $100$ W/cm$^2$). The total power for a $30 \times 30 \mu m^2$ device is in the order of $1mW$. Figures 5.19 and 5.20 show the electron and hole concentrations respectively. The AlGaAs/GaAs heterojunctions can be easily distinguished by the dip in the concentrations as a result of the band discontinuity between larger and smaller band-gap materials. Examining the dynamic behavior of the carriers we observe that the electron concentration recovers back to its steady state condition right after the pulse has been terminated. The hole concentration however slowly regains its steady state value due to its slower
Figure 5.19: Simulated variation of electron concentration in an AlGaAs/GaAs p-i-n photodiode as a function of time and position.

mobility. The transient variation of local electrostatic potential is depicted in Fig. 5.21. The ‘valley’ perceived on the top of the potential profile represents the bias voltage decrease of the detector because of the voltage drop across the resistor due to the photocurrent. Since the electric field in the device also decreases during the optical generation, this lowers the drift component of the carrier transport. This example is clearly a large signal case since the bias conditions are significantly altered by the optical generation.

To demonstrate the dependency of the carrier response times to the electric field even more distinctly, another large signal simulation run in which the photodetector is saturated by a 200 ps pulse was performed. Figure 5.22 shows the impact of such a pulse excitation of 100 W/cm² on the electrostatic potential. The transient bias drop is very significant $25 kT/q$ and in fact, the drop in the detector bias is greater than the external applied voltage. This results in switching from reverse to
forward biasing via the optical pulse. One must notice that once the optical pulse is terminated the electrostatic potential slowly recovers to its steady state value. Fig. 5.23 shows the generation of the hole carriers. The photo-generated carrier concentration is almost equivalent to the doping concentration of the p-GaAs cap layer. Even though the optical pulse is terminated at 0.35 ns the hole concentration still remain stationary in the absorption region for another 0.1 ns. This is a result of the vanishing drift component of the carriers. Once the electrostatic potential (hence the electric field) regains it significance (≈ 0.5 ns) the hole concentration decreases. Even the electrons which have large mobilities at low fields (fig. 5.24) are kept stationary in the absorption region due to the nearly complete quenching of internal electrical field during the pulse.

These examples clearly indicate the importance of large signal analysis since it drastically effects the device speed. Fig 5.25 displays the broadening photocurrent
responses for increasing optical generation rates. The initial photogeneration (the generation rate in which the smallest photocurrent in Fig. 5.25 is obtained) is increased by factors of 5, 10, and 20. When 5 times the generation rate of the initial pulse is applied, the photocurrent response does not increase linearly and pulse broadening is observed. Therefore, we conclude that the generation rate is sufficient enough to saturate the diode. The ringing seen in the photocurrent responses for increasing photogeneration is due to numerical instability in the calculation of the displacement current components. This instability may be avoided by reducing the time step size during the photogeneration process. For example, similar instabilities were observed when obtaining Fig. 5.7. This effect was eliminated by reducing the time step. Since we are more interested in the large signal capability of the model and overall form of the photocurrent response, the time step reduction for the simulation run was not attempted.

Figure 5.21: Simulated variation of electrostatic potential of an AlGaAs/GaAs p-i-n photodiode as a function of time and position.
Figure 5.22: Simulated variation of electrostatic potential of an AlGaAs/GaAs p-i-n photodiode as a function of time and position for a 200 ps duration optical pulse.

5.5 Integrated photoreceiver

As discussed in chapters 2 and 3, simulating optoelectronic devices in a circuit simulator allows for mixed mode device/circuit simulation. This capability is demonstrated with the following example of a simple photoreceiver circuit consisting of a GaAs p-i-n photodiode and a single transistor BJT amplifier circuit (Figure 5.26). The supply voltage $V_{cc}$ is 2.5 V and the bias circuit is adjusted to reverse bias the photodiode and saturate the BJT. In this example, the detailed device level model of the photodiode is combined with the circuit-level lumped models for the rest of the photoreceiver circuit. Since the device-level simulation of the p-i-n photodiode is carried out using the circuit simulator, the link between the device- and circuit-level simulation results is automatically established. The time-dependent carrier concentrations and the local potential yields the terminal current of the simulated device. The transient current-
Figure 5.23: Simulated variation of hole concentration in an AlGaAs/GaAs p-i-n photodiode as a function of time and position for a 200 ps duration optical pulse.

Voltage characteristics of the photodiode is combined with the external circuit and the voltage boundary conditions are evaluated at every time point by solving the entire OEIC self-consistently. This self-consistent solution yields the true large signal response of the device and the OEIC. Figure 5.27 shows the simulated photocurrent of the photodetector and the variation of the detector bias under a 50 ps pulse when approximately 0.5 mW is absorbed in the middle of the depletion region.

Figure 5.28 shows the transient variation of the base current and collector voltage of the transistor in the photoreceiver circuit. Transient variation of base current shows the reversal of bias current under optical pulse. The variation of the collector potential from saturation to nearly to the supply voltage demonstrates the large signal capability of the model. Note that all the circuit variables are evaluated simultaneously with the local device variables. Therefore the device performance can be tested in realistic bias conditions. Besides this mixed-mode device/circuit simulation
Figure 5.24: Simulated variation of electron concentration in an AlGaAs/GaAs p-i-n photodiode as a function of time and position for a 200 ps duration optical pulse.

capability enables the circuit designer to easily and accurately investigate the effects of various subtle changes in device geometry and/or material parameters upon circuit performance. While this task would be difficult or impossible using circuit-level device models only.
Figure 5.25: Photodiode current under increasing optical illumination
Figure 5.26: The simulated photoreceiver circuit composed of a p-i-n diode and a single transistor amplifier. The circuit parameters are $R_1 = 2k\Omega$, $R_2 = 1k\Omega$, $R_c = 100\Omega$, $C = 70\ fF$, and relevant parameters in the SPICE model for BJT are $\beta_f = 500$, $\Pi_i = 10^{-14}$, $\Delta C_{je} = 30\ fF$, $\Delta C_{jc} = 30\ fF$, $\Delta C_{je} = 20\ fF$, $\gamma_c = 10\ Omega$, $\gamma_b = 50\ Omega$ and $\tau_f = 0.01\ ns$. 
Figure 5.27: The transient response of the photoreceiver circuit obtained by mixed-mode device/circuit simulation. Top: diode current (solid line) and photogeneration (dashed line) in arbitrary units. Bottom: transient variation of the photodiode terminal potential.
Figure 5.28: The transient response of the photoreceiver circuit obtained by mixed-mode device/circuit simulation. Transient variation of base current (top) shows the reversal of bias current when the photodiode is illuminated. The variation of the collector potential from saturation to nearly to the supply voltage demonstrates the large signal capability of the model.
Chapter 6

Conclusions

In this thesis, a novel approach has been presented for incorporating the transient solution of one-dimensional semiconductor equations within a general circuit simulation environment. This approach allows simple representation of localized carrier transport models of simulated devices through equivalent circuit elements such as voltage controlled current sources and capacitors. The availability of local photo-generation models at every grid point enables simple simulation of various optoelectronic devices.

The model was utilized to calculate results supporting the theory of the RCE p-i-n and compared it to its conventional counterpart. A three-fold enhancement in the bandwidth-quantum efficiency product was predicted. Large signal capability was demonstrated via the simulation of a heterojunction photodiode.

As the device-level simulation is carried out by the circuit simulator using an equivalent circuit representation, this approach also lends itself to mixed-mode simulation of devices and circuits by a single simulation tool. The mixed-mode simulation capability of the new approach has been tested for the time-domain analysis of a photoreceiver circuit. It has been shown that the transient characteristics of one-dimensional device structures operating within a realistic circuit environment can be simulated without requiring dedicated software tools. The device-level modeling approach introduced in chapters 3 and 4 can easily be extended for mixed two-dimensional device and circuit simulation. The device simulation model presented in this paper has been implemented and tested in SPICE3F2 and also two SPICE-like general-purpose circuit simulation programs. The implementation of the model requires no modifications within the internal structure of the circuit simulator. Thus, it can be implemented quickly and easily on a number of different computation platforms. The versatility of the circuit equivalent representation of the device equations should enable the incorporation of additional differential equations describing other physical phenomena.

The scope of the research presented in this thesis is limited to demonstration of the concept and various capabilities of the described model. Future work may cover experimental verifications of the predictions on specific device performances as well as development of this novel technique into a simulation tool.
Bibliography


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